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(54) Title: METHOD AND APPARATUS FOR ACQUIRING WIDE–BAND PSEUDORANDOM NOISE ENCODED WAVEFORMS

(57) Abstract

The method and apparatus of the present invention is directed to architecture for signal processing, such as for performing analog–to–digital (180) and digital–to–analog (112) conversions, in which the source signal is decomposed into subband signals by an analysis filter (104), processed, and the processed subband signals (228) combined to form a reconstructed signal that is representative of the source signal (102).

## METHOD AND APPARATUS FOR ACQUIRING WIDE-BAND
## PSEUDORANDOM NOISE ENCODED WAVEFORMS

### FIELD OF THE INVENTION

5      The present invention relates generally to a method
and apparatus for acquiring wide-band random and
pseudorandom noise encoded waveforms and specifically to a
method and apparatus for acquiring wide-band signals,
including deterministic signals, random signals and
10     pseudorandom noise encoded waveforms that divides the
waveform into a plurality of subbands prior to signal
processing thereof.

### BACKGROUND

15     Analog-to-digital converters are devices that convert
real world analog signals into a digital representation or
code which a computer can thereafter analyze and
manipulate. Analog signals represent information by means
of continuously variable physical quantities while digital
20     signals represent information by means of differing
discrete physical property states. Converters divide the
full range of the analog signal into a finite number of
levels, called quantization levels, and assigns to each
level a digital code. The total number of quantization
25     levels used by the converter is an indication of its
fidelity and is measured in terms of bits. For example, an
8-bit converter uses $2^8$ or 256 levels, while a 16-bit
converter uses $2^{16}$ or 65536 levels.

During the conversion process, the converter
30     determines the quantization level that is closest to the
amplitude of the analog signal at that time and outputs the
digital code that represents the selected quantization
level. The rate at which the output is created indicates
the speed of the converter and is measured in terms of
35     samples per second (sps) or frequency in Hertz (Hz). As
will be appreciated, a larger number of bits and therefore

The signal can be in any suitable form such as electromagnetic radiation, acoustic, electrical and optical.

In one embodiment, the method includes the following steps:

(a) decomposing the analog or digital signal into a plurality of signal segments (i.e., subband signals), each signal segment having a signal segment bandwidth that is less than the signal bandwidth;

(b) processing each of the signal segments to form a plurality of processed signal segments; and

(c) combining the processed signal segments into a composite signal that is digital when the signal is analog and analog when the signal is digital. As will be appreciated, the sum of the plurality of signal bandwidths is approximately equivalent to the signal bandwidth. The means for processing the signal segments can include any number of operations, including filtering, analog-to-digital or digital-to-analog conversion, signal modulation and/or demodulation, object tracking, RAKE processing, beamforming, null steering, correlation, interference-suppression and matched subspace filtering.

In a particularly preferred application, the signal processing step (b) includes either analog-to-digital or digital-to-analog conversions. The use of signal segments rather than the entire signal for such conversions permits the use of a lower sampling rate to retain substantially all of the information present in the source signal. According to the Bandpass Sampling Theorem, the sampling frequency of the source signal should be at least twice the bandwidth of the source signal to maintain a high fidelity. The ability to use a lower sampling frequency for each of the signal segments while maintaining a high fidelity permits the use of a converter for each signal segment that is operating at a relatively slow rate. Accordingly, a

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In another configuration, noise components in each of the signal segments can be removed prior to signal analysis or conversion in the processing step. The removal of noise prior to analog-to-digital conversion can provide

5    significant additional reductions in computational requirements.

In yet another configuration, a coded signal is acquired rapidly using the above-referenced invention. In the processing step, the signal segments are correlated

10   with a corresponding plurality of replicated signals to provide a corresponding plurality of correlation functions defining a plurality of peaks; an amplitude, time delay, and phase delay are determined for at least a portion of the plurality of peaks; and at least a portion of the

15   signal defined by the signal segments is realigned and scaled based on one or more of the amplitude, time delay, and phase delay for each of the plurality of peaks.

In another embodiment, a method is provided for reducing noise in a signal expressed by a random or

20   pseudorandom waveform. The method includes the steps of decomposing the signal into a plurality of signal segments and removing a noise component from each of the signal segments to form a corresponding plurality of processed signal segments. The means for decomposing the signal can

25   be any of the devices noted above, and the means for removing the noise component includes a noise reducing quantizer, noise filters and rank reduction. Signal reconstruction may or may not be used to process further the processed signal segments. This embodiment is

30   particularly useful in acquiring analog signals.

In yet a further embodiment, a method is provided for combining a plurality of signal segments (which may or may not be produced by analysis filters). In the method, synthesis filtering is performed on each of the plurality

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oblique projection filtration (e.g., as described in copending U.S. Patent Application Serial No. 08/916,884 filed August 22, 1997, entitled "RAKE Receiver For Spread Spectrum Signal Demodulation," which is incorporated herein
5  fully by reference).


BRIEF DESCRIPTION OF THE DRAWINGS
Figure 1 depicts a first embodiment of the present invention;
10  Figure 2 depicts an analog signal;
Figure 3 depicts the analog signal of Figure 2 divided up into a plurality of signal segments;
Figure 4 depicts the first embodiment including decimation;
15  Figures 5A and 5B depict noble identities;
Figure 6 depicts a polyphase filter representation;
Figure 7 depicts a polyphase filter representation with noble identities;
Figure 8 depicts another embodiment of the present
20  invention;
Figure 9 depicts the quantization process of the quantizers in Figure 8;
Figure 10 depicts a subband digital transmitter;
Figure 11 depicts a subband analog transmitter;
25  Figure 12 depicts a subband receiver;
Figure 13 depicts rank reduction for noise filtering;
Figure 14 depicts another embodiment of the present invention;
Figure 15 depicts another embodiment of the present
30  invention;
Figure 16 depicts RAKE processing;

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The analysis and synthesis filters 44a-n and 60a-n can be in any of a number of configurations provided that the filters pass only discrete, or at most only slightly overlapping, portions of the frequency domain of the signal

5   40. It is preferred that the frequency bands of the subband signals overlap as little as possible. Preferably, no more than about 5% and more preferably no more than about 1% of the frequency bands of adjacent subband signals overlap.

10  The filters can be analog or digital depending on the type of signal 40 or the processed signal segments 56a-n. Examples of suitable analog analysis and synthesis filters include a suitably configured bandpass filter formed by one or more low pass filters, one or more high pass filters, a

15  combination of band reject and low pass filters, a combination of band reject and high pass filters, or one or more band reject filters. Digital analysis and synthesis filters are typically defined by software architecture that provides the desired filter response.

20  In a preferred configuration shown in Figure 4, the signal 40 is decomposed by the analysis filter bank 46 (which includes analog or digital analysis filters $H_k(z)$ 44a-n) into subband signals which are each sampled by a downsampler 64a-n performing an M-fold decimation (i.e.,

25  taking every $M^{th}$ sample), and the sampled subband signals are further sampled after signal processing by an up-sampler 68a-n (and/or expander (which fills in L-1 zeros in between each sample)) and the further sampled subband signals are combined by a synthesis filter bank 62 (that

30  includes analog or digital synthesis filters $G_k(z)$ 60a-n). The sampled subband signals, denoted by $x_0(n)$, $x_1(n)$, . . . $x_{m-1}(n)$, are the outputs of the N-band analysis filter bank and the inputs to the N-band synthesis filter bank. As a result of decimation, the subband signals are 1/N the rate

35  of the input rate of the signal 40.

$$G^T(z) = G_o^T(z)U(z)$$

where $G_o^T(z)$ is the minimum phase equivalent of $G^T(z)$.

In a preferred embodiment, the rational transfer matrices of the analysis and/or synthesis filters are mathematically expressed in a polyphase filter representation. Exemplary equations defining the decomposition of the signal 40 by the analysis filters 44a-n include the following:

$$H(z) = \sum_{l=0}^{M-1} z^1 E_l(z^M)$$

where

M is the number of subbands (which is the same as the number of analysis filters in the analysis filter bank;

$l$ is the subband designation);

$$E_l(z^M) = \sum_{n=-\infty}^{\infty} e_l(n) z^{-n}$$

$$e_l(n) = h(Mn+l), 0 \leq l \leq M-1$$

(known as a Type 1 polyphase filter representation) and

$$H(z) = \sum_{l=0}^{M-1} z^{-(M-1-l)} R_l(z^u)$$

where

$$R_l(z^M) = E_{M-1-l}(z)$$

(known as Type 2 polyphase filter representation). As will be appreciated, other techniques exist for expressing a rational transfer matrix defining a filter system including impluse response and filter description.

Noble identities can be used to losslessly move the decimators to the left of the analysis filters and the L-fold up-sampler and/or expander to the right of the synthesis filters. In this manner, the analysis and synthesis filters operate on lower rate data, thereby

convert the subband signals 48a-n to digital form before further processing 82 (e.g., correlation for encoded subband signals, subband signal digital beamforming in multiple antenna systems, etc.) and/or synthesis of the digital subband signals 78a-n is performed. As noted above, the subband signals 48a-n are preferably sampled by each of the decimators or downsamplers 64a-n at a rate of at least about twice the bandwidth of the corresponding subband signal 48a-n to maintain fidelity. As shown in Fig. 9, each quantizer, or analog-to-digital converter, 74a-n determines the digital word or representation 90a-n that corresponds to the bin 86a-n having boundaries capturing the amplitude of the analog subband signal at that time and outputs the digital word or representation that represents the selected quantization level assigned to the respective bin. The digital subband signals 78a-n are converted 94a-n from radio frequency (RF) to base band frequency and optionally subjected to further signal processing 60. The processed subband signals 98 are formed into a digital composite signal 102 by the synthesis filter bank 60.

To provide increased accuracy, noise rejecting quantizers can be utilized as the quantizers 74a-n. As will be appreciated, a noise rejecting quantizer assigns more bits to the portions of the subband signal having less noise (and therefore more signal) and fewer bits to the noisy portion. This selective assignment is accomplished by adaptively moving the bin boundaries so as to narrow the bin width (thereby increasing quantization fidelity. An example of a design equation for a Lloyd-Max noise rejecting quantizer is as follows:

$$t_k = \frac{x_{k-1} + x_k}{2} + \frac{\delta^2(x_k) - \delta^2(x_{k-1})}{2(x_k - x_{k-1})} \; ; \; x_k = e_k - \frac{1}{2}\frac{d\delta^2(x_k)}{dx_k}$$

where:

where T is the Wiener filter, $R_{xy}$ is the cross covariance between x and y and $R_{yy}$ is the covariance of y.

$R_{xy}$ and $R_{yy}$ are unknown and require estimation. A number of techniques can be used to estimate $R_{xy}$ and $R_{yy}$, including an adaptive Wiener filter (e.g., using the linear mean squared algorithm), direct estimation, sample matrix inversion and a recursive, adaptive Wiener filter, with a recursive, adaptive Wiener filter being more preferred.

The recursive, adaptive Wiener filter is explained in Thomas, J.K., *Canonical Correlations and Adaptive Subspace Filtering*, Ph.D Dissertation, University of Colorado Boulder, Department of Electrical and Compute Engineering, pp.1-110, June 1996. which is incorporated herein by reference in its entirety. In a recursive, adaptive Wiener filter assume $\hat{T}_M$ denotes the filter when M measurements of X and Y are used. Then $\hat{T}_M$ is the adaptive Wiener filter

$$\hat{T}_M = X_M Y_M * (Y_M Y_M *)^{-1} = \hat{R}_{xy} \hat{R}_{yy}^{-1},$$

$$X_M = \begin{bmatrix} x_1 x_2 \cdots x_M \end{bmatrix}; \quad X_{M+1} = \begin{bmatrix} X_M x \end{bmatrix}$$

$$Y_M = \begin{bmatrix} y_1 y_2 \cdots y_M \end{bmatrix}; \quad Y_{M+1} = \begin{bmatrix} Y_M y \end{bmatrix}$$

If another measurement of x and y is taken, and one more column is added to $X_M$ and $Y_M$ to build $\hat{T}_{M+1}$ :

$$\hat{T}_{M+1} = X_M Y_M * \hat{R}_{M+1}^{-1} + xy * \hat{R}_{M+1}^{-1}$$

The estimate of $_{M+1}$ is $\hat{X}_{M+1}$

$$\hat{X}_{M+1} = \hat{T}_{M+1} Y_{M+1}$$

Using the estimate of $X_{M+1}$, one can read off $\hat{x}_{M+1}$, which is the estimate of x:

$$\hat{x}_{M+1} = \frac{1}{1+r^2} \tilde{x}_M + \frac{r^2}{1+r^2}$$

where $r^2 = y * \hat{R}_M^{-1} y$ and $\tilde{x}_{M+1} = \hat{T}_M y$.

Based on the above, when one observes y, the best estimate of the unknown x is $\tilde{x}$, with corresponding estimation error $\tilde{E}_{M+1}$ and covariance $\tilde{Q}_{M+1}$. If the unknown x becomes available after a delay, then $\tilde{x}_{M+1}$ can be updated to $\hat{x}_{M+1}$ with error

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In another embodiment, the source signal 40 is digital and the analysis filters are therefore digital, signal processing is performed by a digital-to-analog converter, and the synthesis filters are analog. Figure 10 depicts a
5   subband digital transmitter according to this embodiment. The signal 100 is in digital format and is transmitted to a bank of analysis filters 104a-n to form a plurality of digital subband signals 108a-n; the digital subband signals 108a-n are processed by digital-to-analog converters 112a-n
10   to form analog subband signals 116a-n; the analog subband signals 116a-n are amplified by amplifiers 120a-n to form amplified subband signals 124a-n; and the amplified subband signals 124a-n transmitted via antennas 128a-n.

In another embodiment shown in Figure 11, a subband
15   analog transmitter is depicted where the signal 140 is analog and not digital. The signal 140 is decomposed into a plurality of analog subband signals 144a-n by analog analysis filters 148a-n and the analog subband signals 144a-n amplified by amplifiers 152a-n, and the amplified
20   subband signals transmitted by antennas 156a-n.

In yet another embodiment shown in Figure 12, a subband receiver is depicted that is compatible with the subband analog transmitter of Figure 11. Referring to Figure 12, a plurality of subband signals 160a-n are
25   received by a plurality of antennas 164a-n, the received subband signals 168a-n down converted from radio frequency to baseband frequency by down converters 172a-n; the down converted subband signals 176a-n which are in analog form are converted by quantizers 180a-n from analog to digital
30   format; and the digital subband signals 184a-n combined by synthesis filters 188a-n to form the digital composite signal 192.

In any of the above-described transmitter or receiver embodiments, when the subband signals are encoded waveforms
35   such as Code Division Multiple Access (CDMA) or precision

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correlation function and the noise-free version of the correlation function is significantly reduced.

As shown in Figure 14 to perform the correlation in the subband signals in GPS, CDMA, and other pseudorandom or
5   random waveform applications, the replicated code 208 from the code generator 212 must be passed through an analysis filter bank 216 that is identical to the analysis filter bank 220 used to decompose the signal 224. Because the correlation must be performed for different segments of the
10   replicated code 208, each indexed by some start time, this decomposition is necessary for all trial segments of the replicated code 208. A plurality of subband correlators 228a-n receive both the subband signals 232a-n and the replicated subband signals 236a-n and generate a plurality
15   of subband correlation signals 240a-n. The subband correlation signals 240a-n are provided by the following equation:

$$q_{m,n}^{(i)}(j) = \sum_{k=1}^{N} x_m(k+j)\, p_n^{(i)}(k)$$

20   where:

      q(k)   is the subband correlation signal;

$p_n^{(i)}(k)$ is the component of the $i^{th}$ trial segment of

the P(Y) code in the $n^{th}$ subband;
25

     $x_m(k)$   is the component of the measurement that lies in the $m^{th}$ subband;

     N   is the number of samples over which the correlation is performed.

30   The subband correlation signals 240a-n are upsampled and interpolated by the synthesis filters 244a-n and then squared and combined. The resulting composite signal 248

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function 344 corresponding to the signal 300. The
correlation function 344 is passed to a pre-detector 348 to
determine an estimated transmit time and frequency and an
amplitude and delay for each of the correlation peaks. The

5        estimated transmit time and frequency 352 are provided to
a code generator 356 and the amplitude and time delay 360
associated with each correlation peak are provided to the
RAKE processor 364. The code generator 356 determines a
replicated code 368 corresponding to the signal 300 based

10       on the estimated trial time and frequency. Using the
correlation peak amplitudes and time and/or phase delays,
the RAKE processor 364, as shown in Figure 16, shifts the
input sequence y(k) by the amounts of the multipath time
and/or phase delays and then weights each shifted version

15       by the amplitude of the peak of the correlation function
corresponding to that peak to form a RAKED signal 372
(denoted by $y_R(k)$). The RAKED sequence is commonly defined
by the following mathematical equation:

$$y_R(k) = \frac{1}{\displaystyle\sum_{i=1}^{p} A_i} \sum_{i=1}^{p} A_i e^{-j\phi_i} y(k + t_i)$$

where:

20           p is the number of multipath signals (and therefore the
             number of peaks);
         $A_i$ is the amplitude of the $i^{th}$ peak;
         $t_i$ is the time delay of the $i^{th}$ peak;
         $\phi$ is the phase delay of the $i^{th}$ peak;

25       y(k) is the input sequence into the code correlator.
The RAKED signal 372 and the replicated code 368 are
correlated in a correlator 376 to provide the actual
transmit time and frequency 380 which are then used by a
detector 384 to detect the signal.

The detected signals 438a-n are analyzed by a synthesis filter bank 412a-n to form a composite radar signal 446.

In a variation of the system of Figure 15, a bank of analysis filters and synthesis filters can be implemented both directly before and after the correlation step (not shown) to provide the above-noted reductions in computational requirements.

In another variation of the system of Figure 15, the analysis filters can be relocated before the analog-to-digital converter 316 to form the subband signals before as opposed to after conversion.

In another variation shown of the system of Figure 15 that is depicted in Figure 20, the RAKE processor 364 can account for the relative delays in antenna outputs of the signal 300 (which is a function of the arrangement of the antennas as well as the angular location of the signal source) by summing the antenna outputs without compensating for the relative output delays. The correlation process may result in Nxp peaks, where N is the number of antenna outputs and p is the number of multipath induced peaks. The Np peaks are then used to realign and scale the input data before summation. The RAKE 364 in effect has performed the phase-delay compensation usually done in beam-steering. The advantages of this approach compared to conventional beam steering techniques include that it is independent of antenna array geometries and steering vectors, it does not require iterative searches for directions as in LMS and its variants, and it is computationally very efficient. This approach is discussed in detail in copending application having Serial No. 08/916,884, and filed on August 21, 1997.

While various embodiments of the present invention have been described in detail, it is apparent that modifications and adaptations of those embodiments will occur to those skilled in the art. However, it is to be

What is claimed is:

1.    A method for acquiring a signal having a bandwidth, comprising:

decomposing the signal into a plurality of signal segments, each signal segment having a signal segment bandwidth that is less than the signal bandwidth;

processing each of the signal segments to form a plurality of processed signal segments; and

combining the processed signal segments into a composite signal wherein the signal is one of analog or digital and the composite signal is the other one of analog or digital.

2.    The method of Claim 1, wherein the processing step includes performing analog-to-digital conversion of each of the signal segments.

3.    The method of Claim 1, wherein the processing step includes performing digital-to-analog conversion of each of the signal segments.

4.    The method of Claim 1, wherein the processing step includes removing a noise component from each of the signal segments to form a corresponding plurality of noise reduced signal segments and thereafter converting each of the noise reduced signal segments from one of analog or digital format to the other of analog or digital format.

repeating the assigning, sampling, comparing and selecting steps.

12. The method of Claim 1, wherein the processing step comprises:

correlating the plurality of signal segments with a corresponding plurality of replicated signal segments to provide a corresponding plurality of correlation functions.

13. The method of Claim 12, wherein the processing step comprises:

determining an amplitude, time delay, and phase delay for at least a portion of a plurality of peaks defined by the plurality of correlation functions and

realigning and scaling at least a portion of the signal defined by the signal segments based on one or more of the amplitude, time delay, and phase delay for the at least a portion of the plurality of peaks.

14. An apparatus for acquiring a signal having a signal bandwidth, comprising:

means for receiving a signal in the form pseudorandom or random waveform having a signal bandwidth;

means for decomposing the signal into a plurality of signal segments, each signal segment having a signal segment bandwidth that is less than the signal bandwidth;

means for processing each of the signal segments to form a plurality of processed signal segments; and

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22. A method for reducing noise in a signal having a bandwidth, comprising:

decomposing the signal into a plurality of signal segments, each signal segment having a bandwidth that is less than the bandwidth of the signal and

removing a noise component from each of the signal segments to form a corresponding plurality of processed signal segments.

23. The method of Claim 22, further comprising:

combining each of the processed signal segments to form a composite signal

24. The method of Claim 23, wherein the composite signal has the same bandwidth as the signal.

25. A system for reducing noise in a signal having a bandwidth, comprising:

means for decomposing the signal into a plurality of signal segments, each signal segment having a bandwidth that is less than the bandwidth of the signal and

means for removing a noise component from each of the signal segments to form a corresponding plurality of processed signal segments.

26. The system of Claim 25, further comprising:

means for combining each of the processed signal segments to form a composite signal.

27. The system of Claim 26, wherein the composite signal has the same bandwidth as the signal.

means for receiving each of the plurality of signal segments.

33. The system of Claim 31, further comprising:

means for converting each of the signal segments from

5    an analog format to a digital format.

34. The system of Claim 31, further comprising:

a plurality of analysis filters to decompose a source signal into a plurality of decomposed signal segments;

a plurality of digital-to-analog conversion devices

10   for converting the plurality of decomposed signal segments from digital into analog format to form a corresponding plurality of analog signal segments;

a plurality of amplifiers to form a corresponding plurality of signal segments;

15   a plurality of signal emitters for emitting the plurality of signal segments; and

a plurality of receptors for receiving the plurality of signal segments.

35. The system of Claim 31, further comprising:

20   a plurality of analysis filters to decompose a source signal into a plurality of decomposed signal segments;

a plurality of amplifiers to amplify the decomposed signal segments to form a corresponding plurality of signal segments;

25   a plurality of signal emitters for emitting the plurality of signal segments; and

processing each of the analog signal segments to form a plurality of processed analog signal segments; and

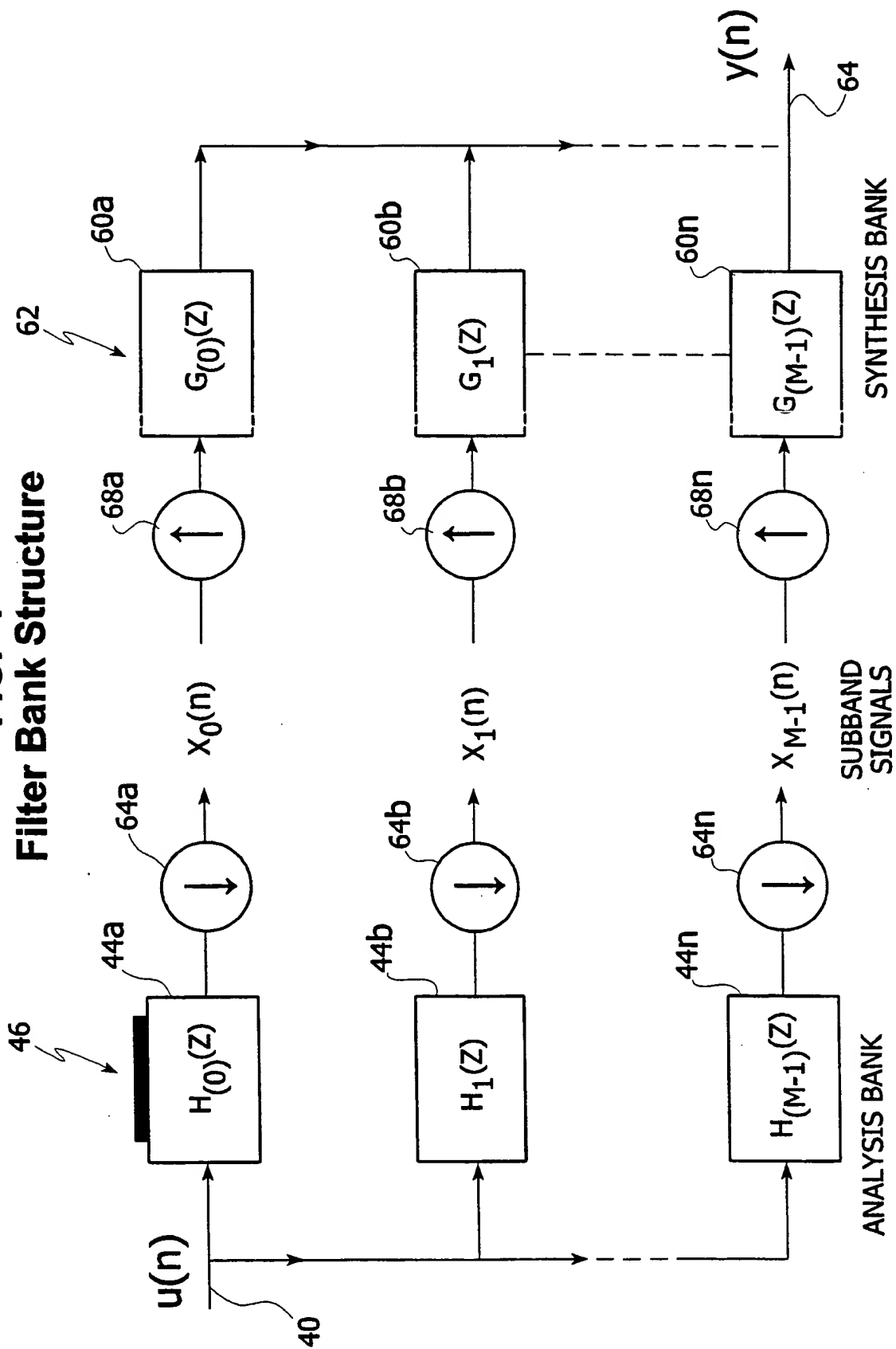combining the processed analog signal segments into a composite signal.
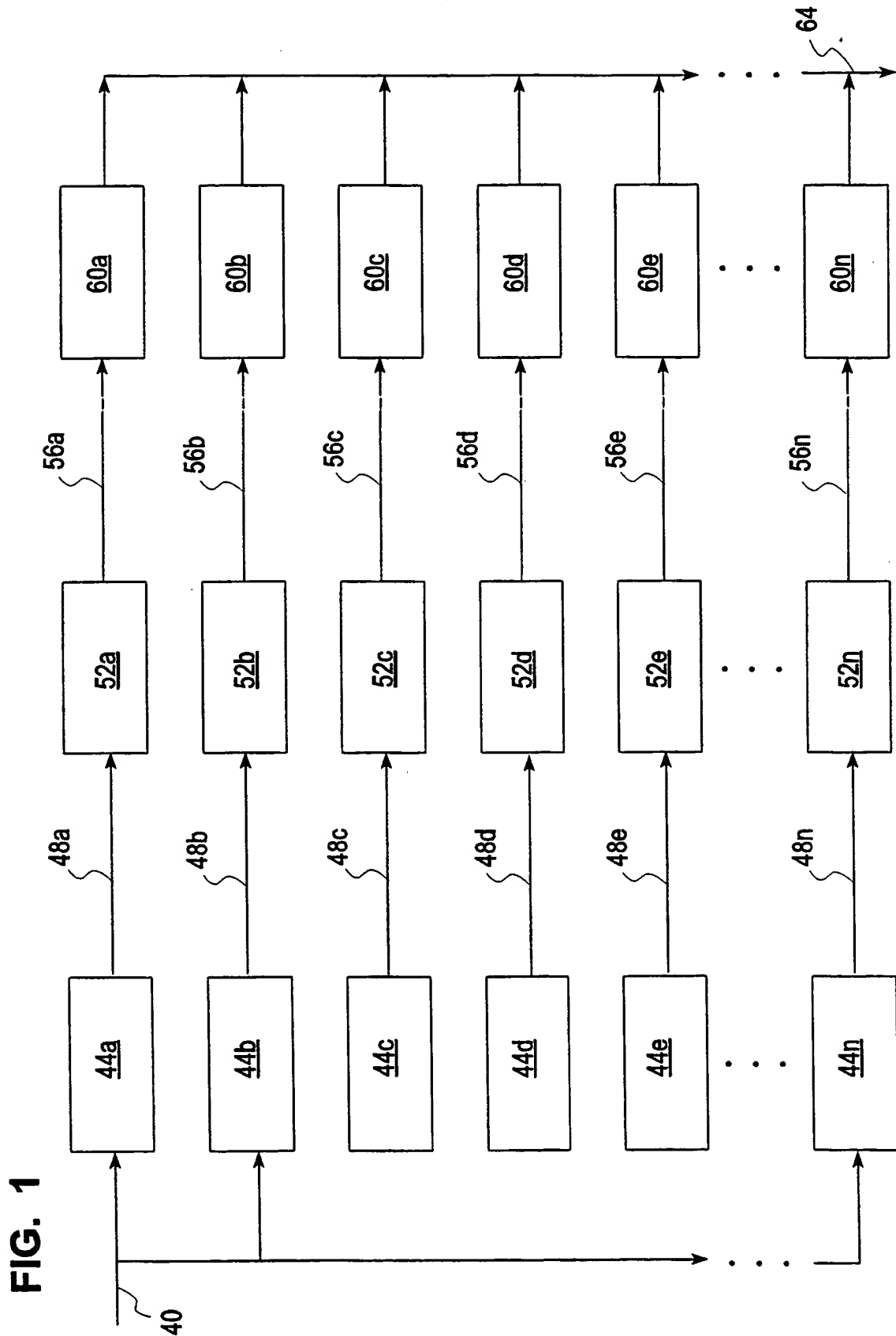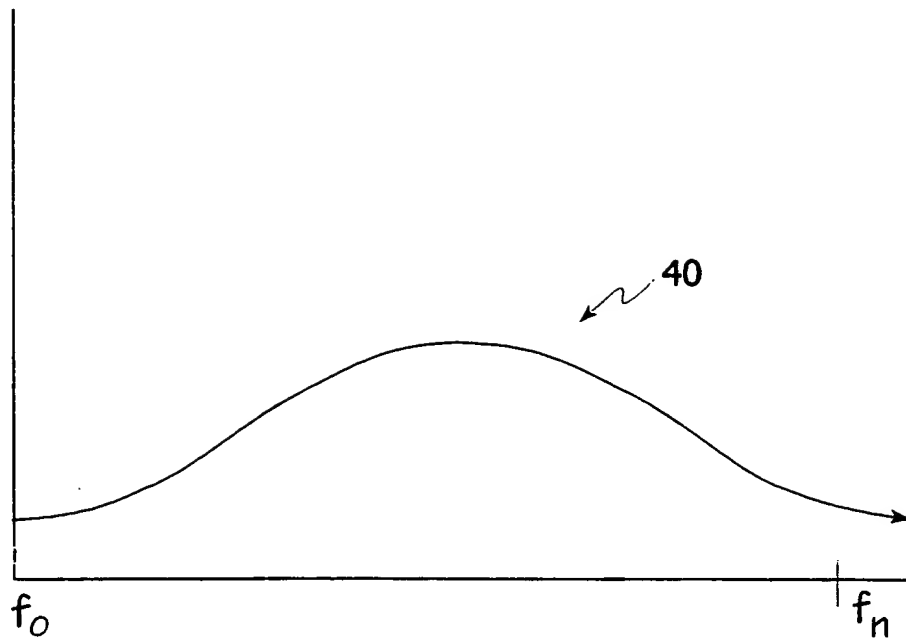
**FIG. 1**

**FIG. 2**
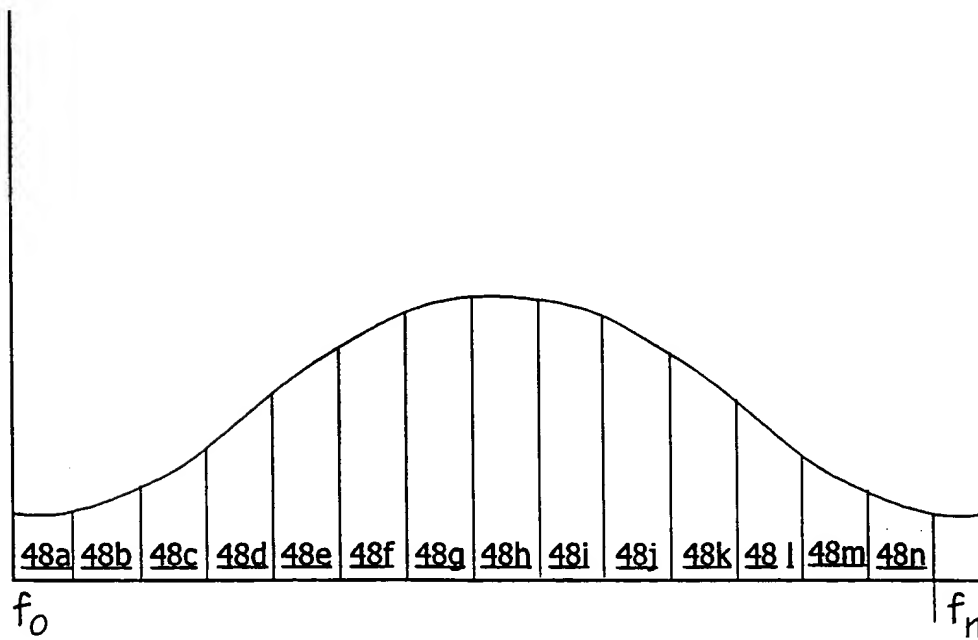


**FIG. 3**

# FIG. 4
## Filter Bank Structure

$$x(n) = u(n)$$

FIG. 5A



$$y(n) = x(n)$$
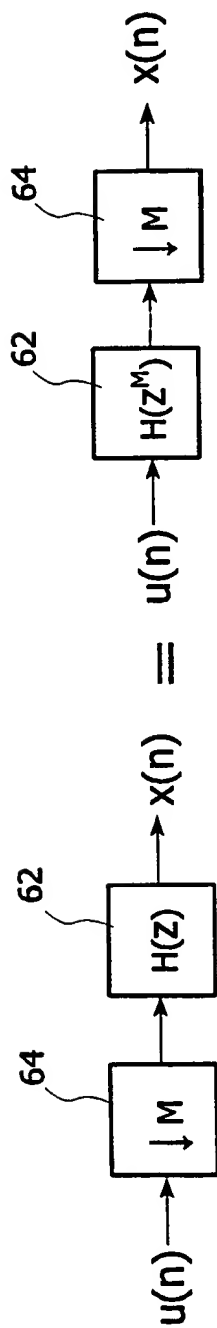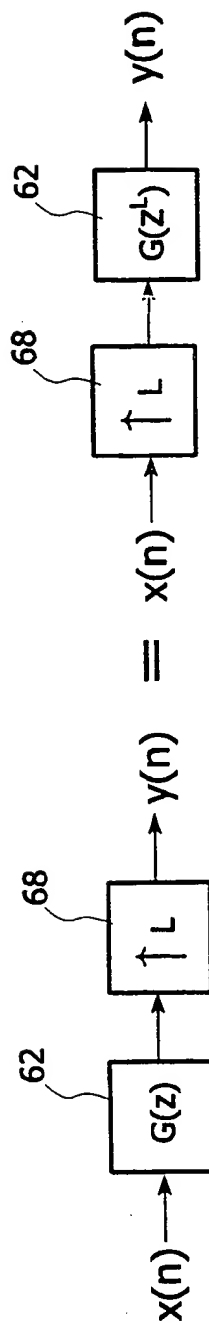
FIG. 5B

NOBLE IDENTITIES I AND II

POLYPHASE AND NOBLE IDENTITIES

FIG. 7



POLYPHASE REPRESENTATION

FIG. 6

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ANTENNA

↓ 40

| BANK OF LOW-NOISE NARROW BANDWIDTH AMPLIFIERS |   44

48d      48c      48b      48a

64n  (←)    LOW RATE SAMPLERS    (←) 64c    (←) 64b    (←) 64a

74n      74c      74b      74a

QUANTIZER      QUANTIZER      QUANTIZER      QUANTIZER

78n      78c      78b      78a

RF TO B BAND      RF TO B BAND      RF TO B BAND      RF TO B BAND

94n      94c      94b      94a

| APPLICATION |   82

↓ 98

| BANK OF SYNTHESIS FILTERS |   60

↓ 102

DIGITAL DATA
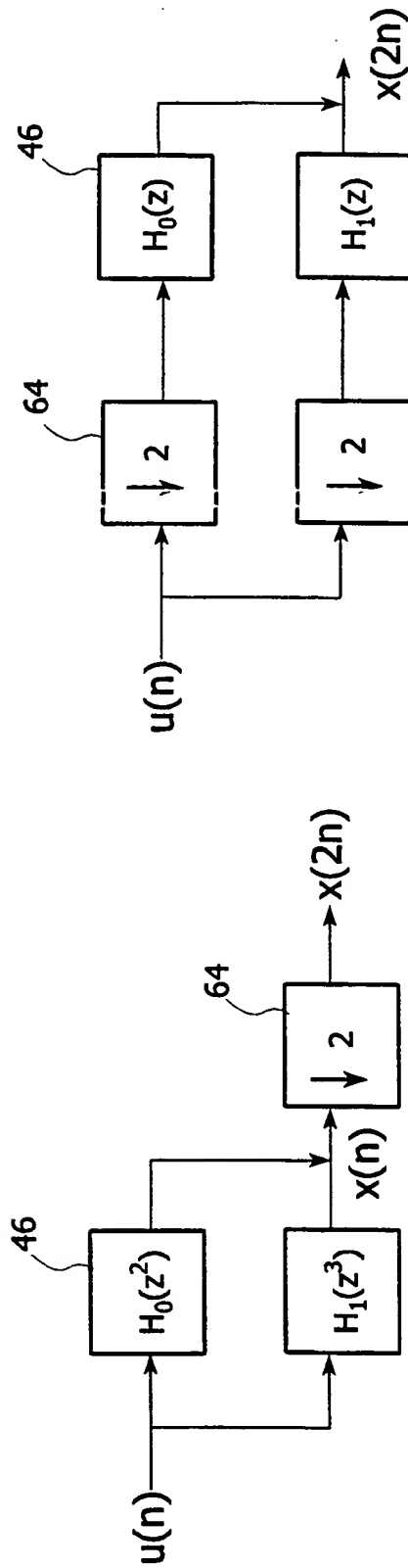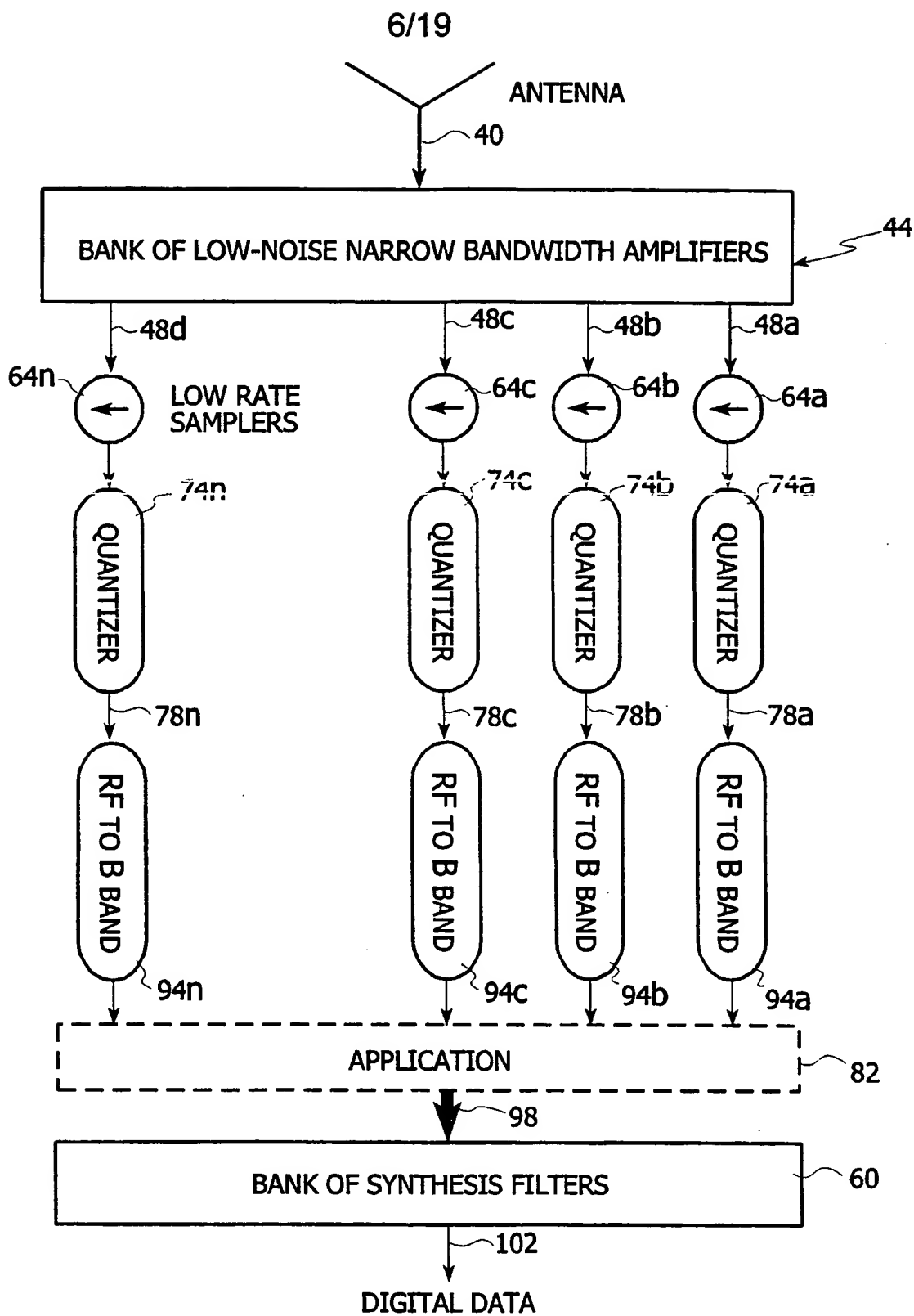
**FIG. 8**
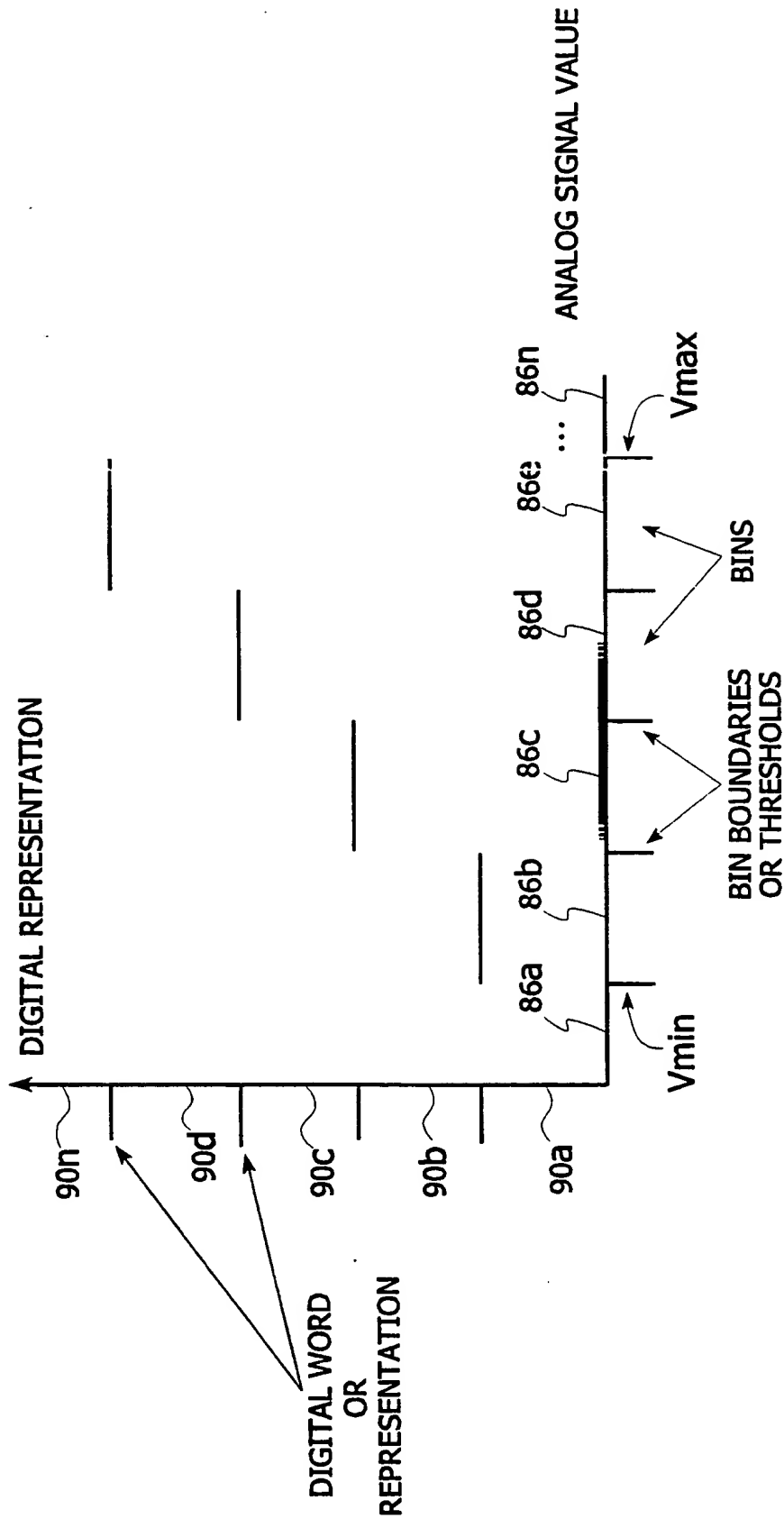
FIG. 9

QUANTIZATION IN AN ADC

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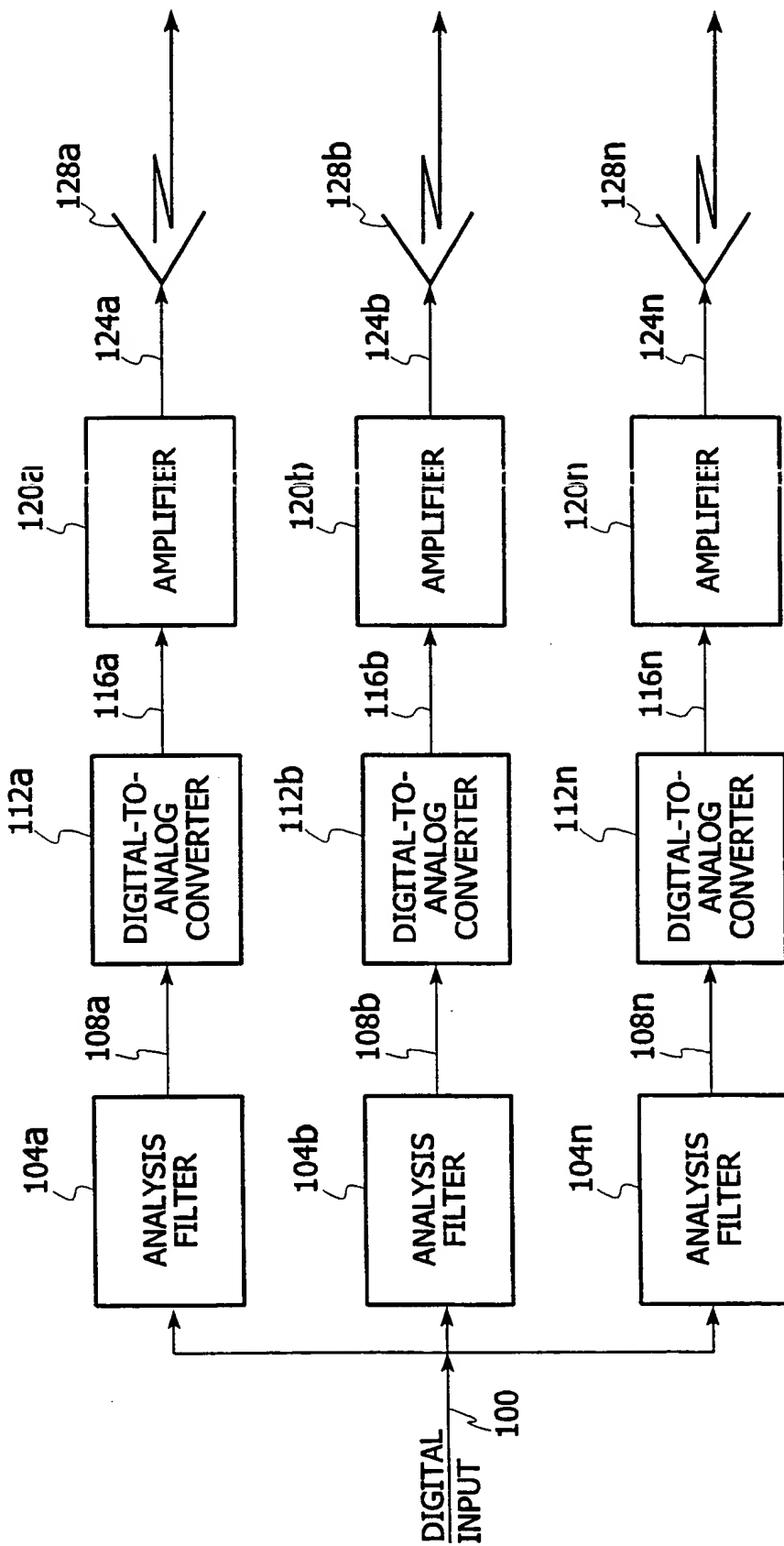


FIG. 10
SUBBAND DIGITAL TRANSMITTER

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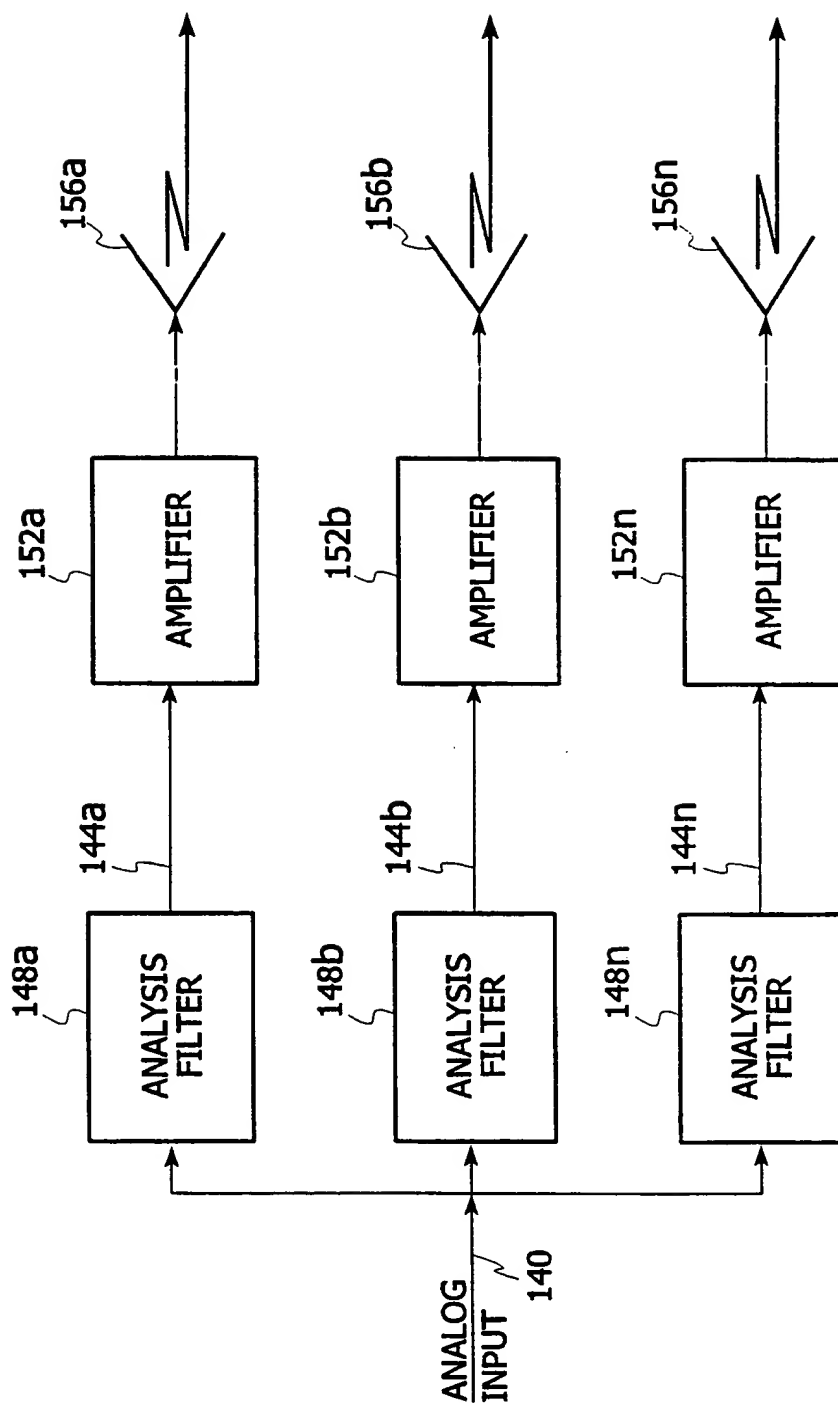


FIG. 11
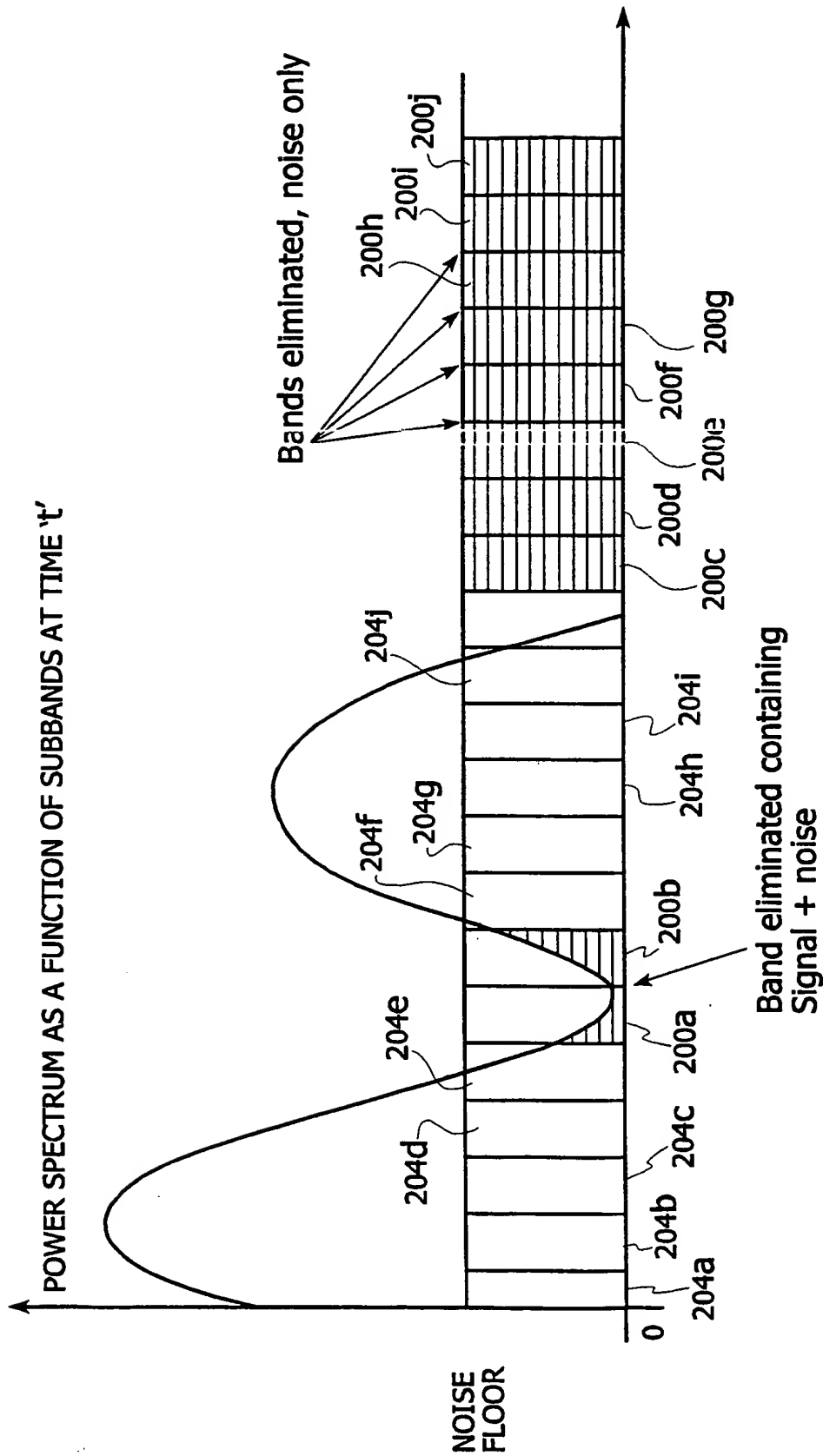SUBBAND ANALOG TRANSMITTER
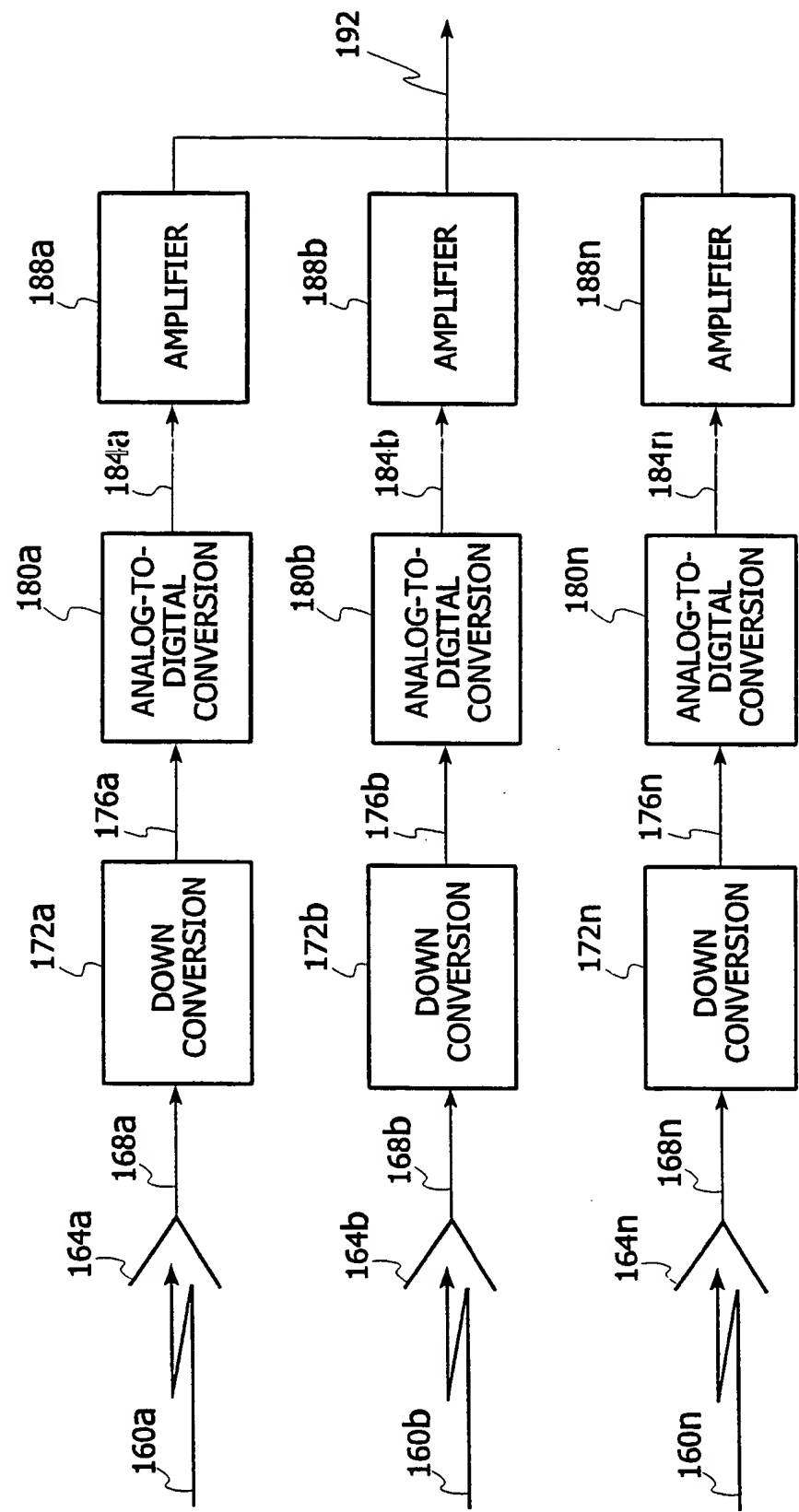
**FIG. 12**
**SUBBAND RECEIVER**

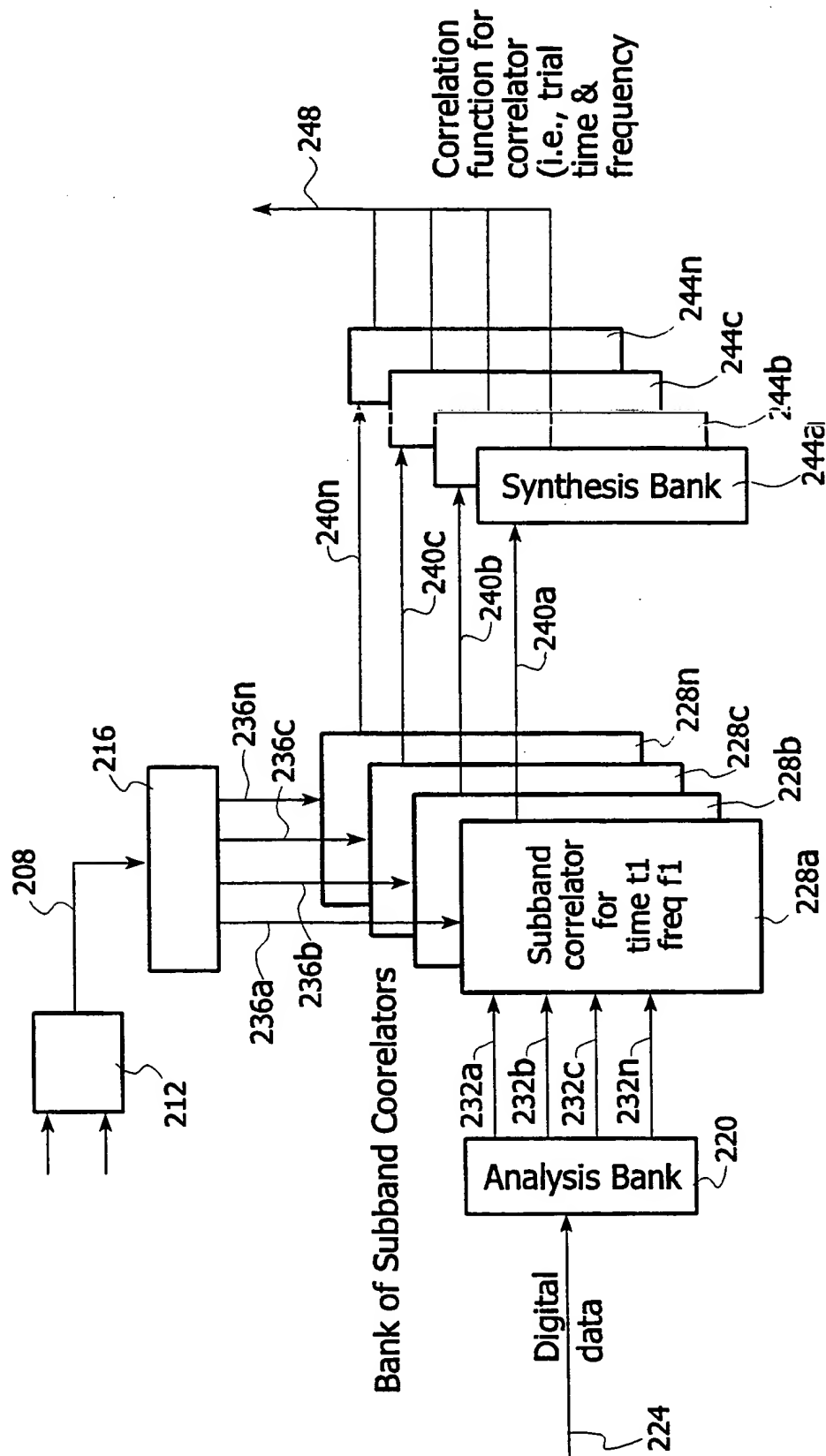## FIG. 13
### RANK-REDUCTION FOR NOISE FILTERING

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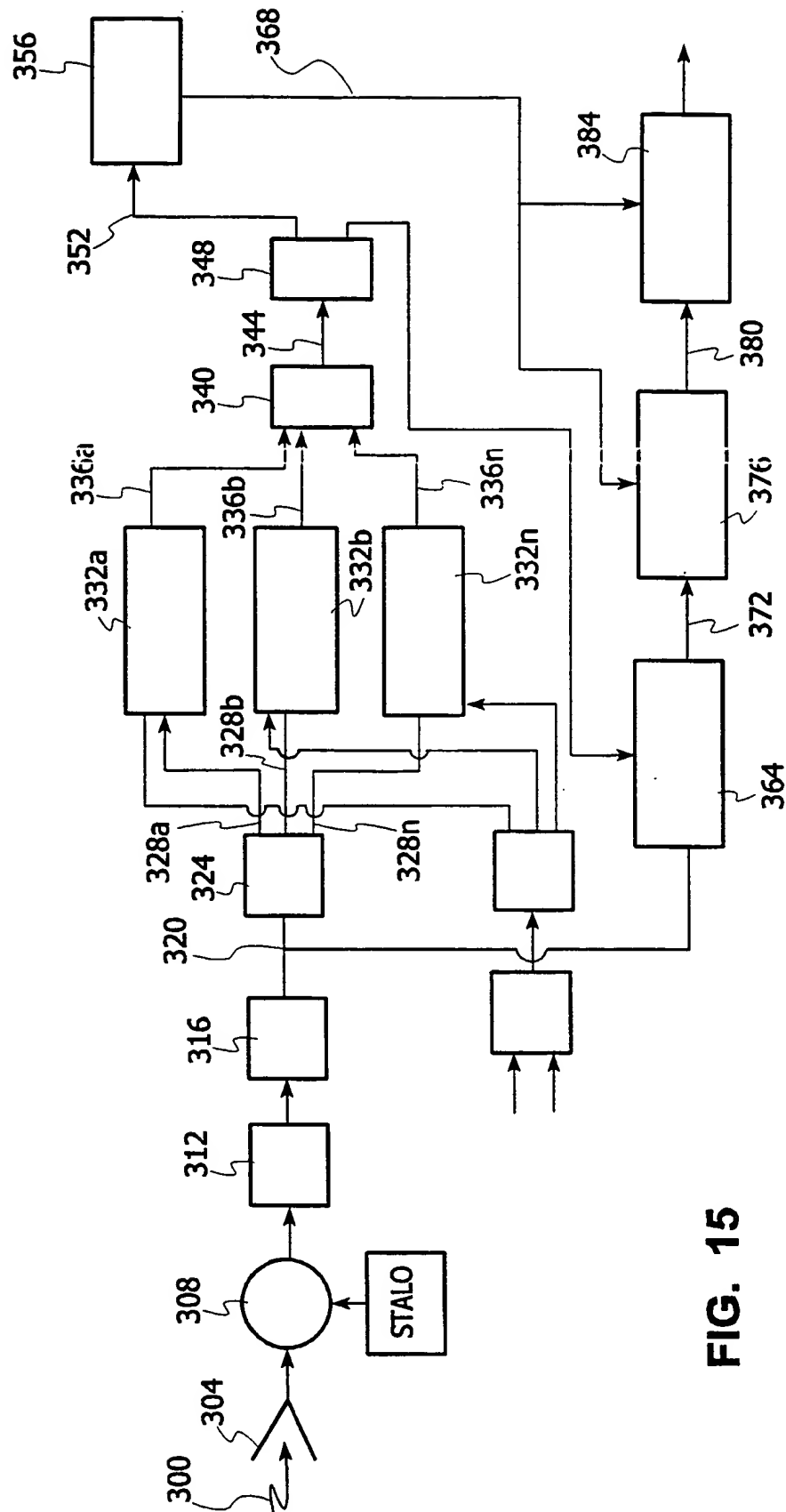
**FIG. 14**
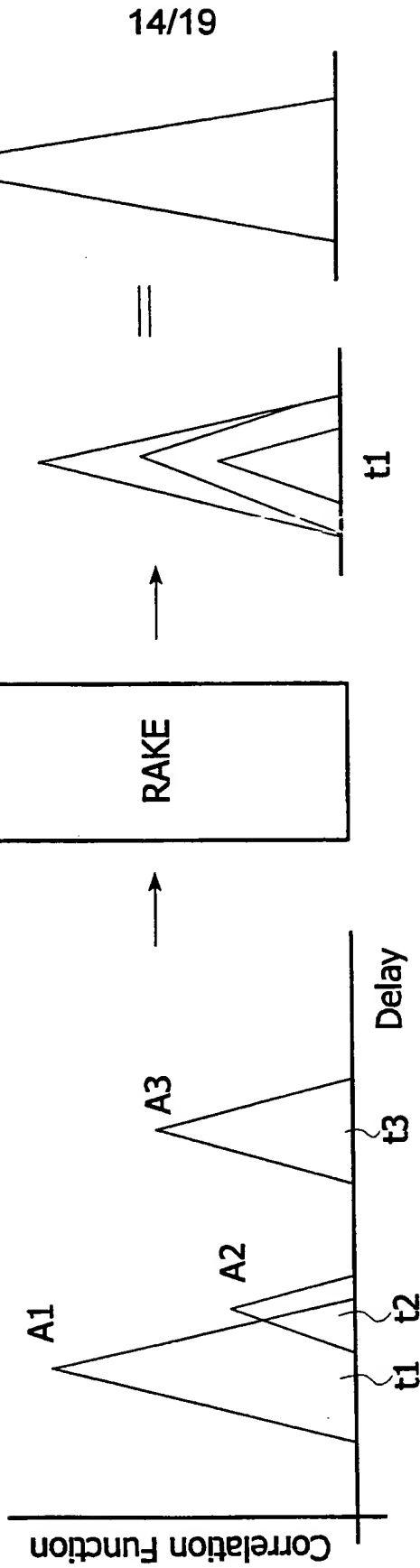
**GPS SIGNAL CORRELATION WITHIN A FILTER BANK**
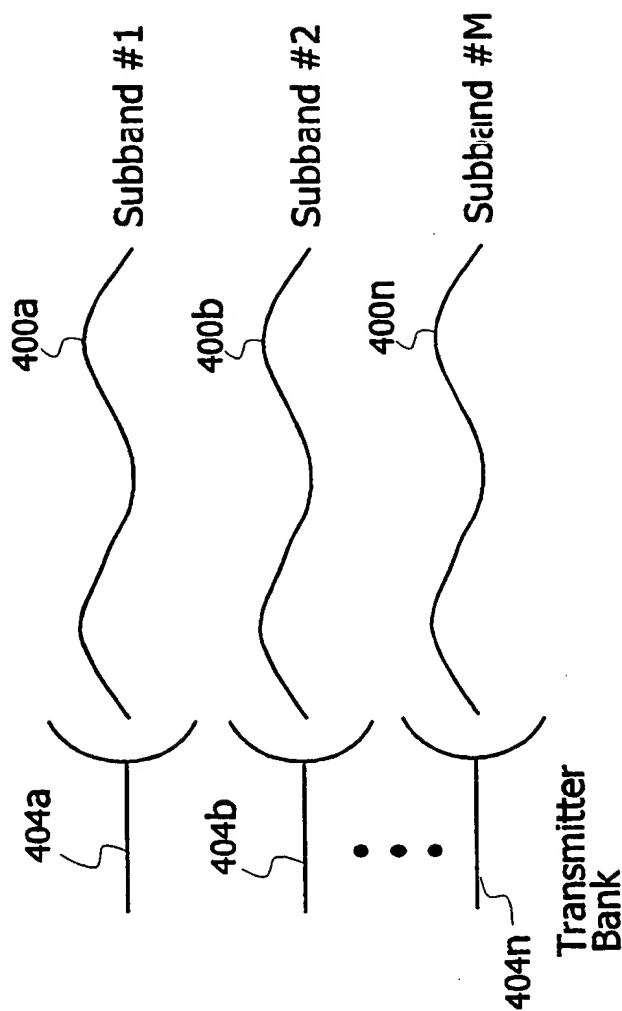
FIG. 15

**FIG. 16**
**RAKE PROCESSING**

Subband #1

400a

404a

Subband #2

400b

404b

Subband #M

400n

404n

Transmitter Bank

• • •

FIG. 17

PROPOSED MULTIPLEXED RADAR TRANSMITTER ARCHITECTURE

Digital Radar Data

Synthesis Filter Bank 412

438a | 438b | 438n

Detector #1 434a | Detector #2 434b | ... | Detector #M 434n

430a | 430b | 430n

Quantizer 426a | Quantizer 426b | Quantizer 426n

A/D Converters in Subbands

422a | 422b | 422n

416a | 416b | 416n

Bank of Low Pass Filters 414

410

Receiver

400a | 400b | 400n

Subband #1

Subband #2

. . .

Subband #M
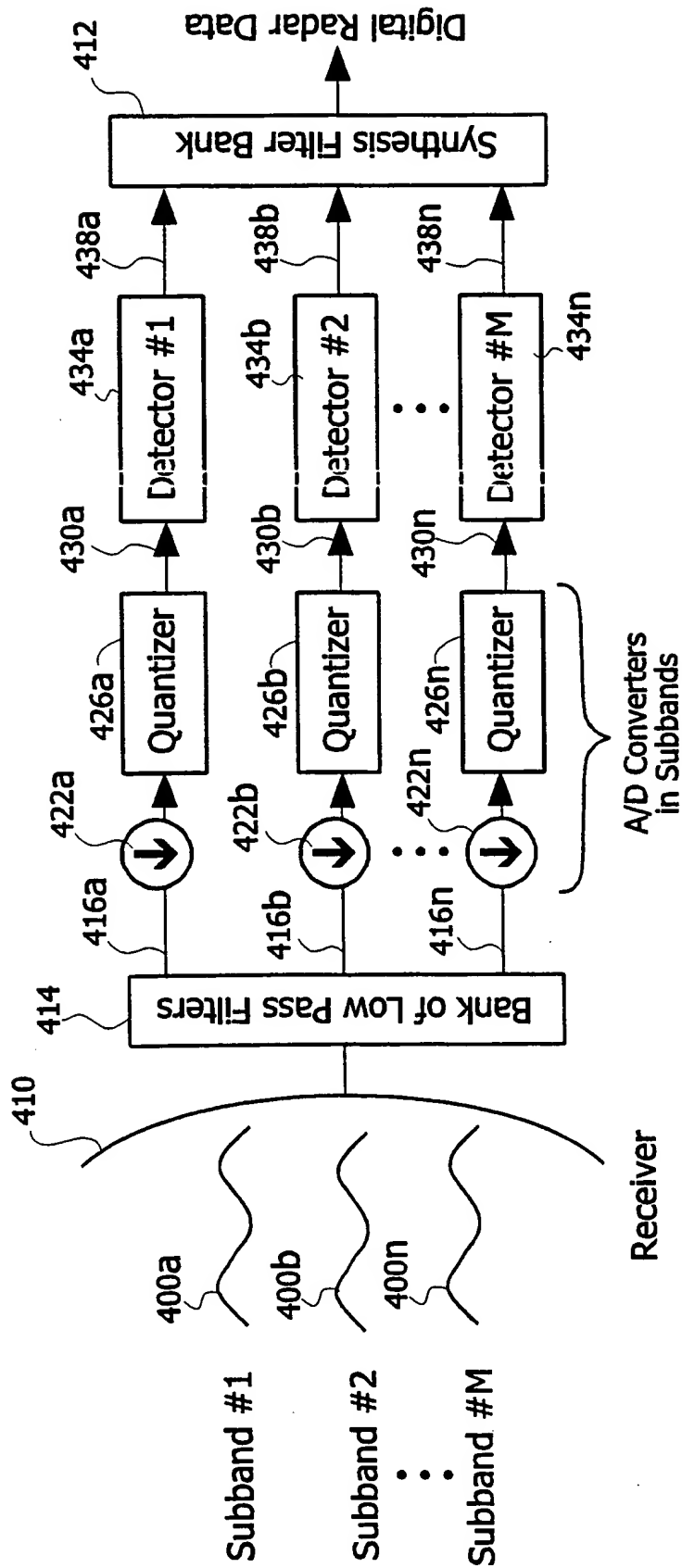
**FIG. 18**

**PROPOSED RADAR RECEIVER ARCHITECTURE**
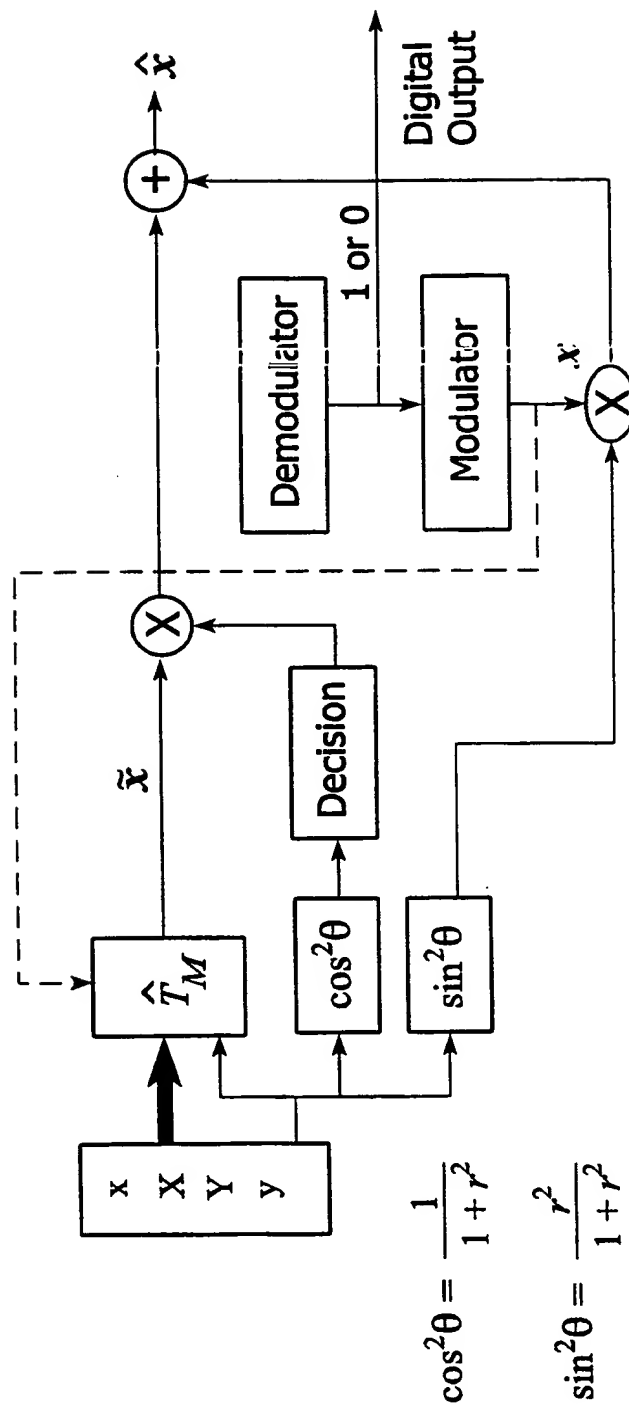
FIG. 19

DIGITAL COMMUNICATION EXAMPLE

$$\cos^2\theta = \frac{1}{1+r^2}$$

$$\sin^2\theta = \frac{r^2}{1+r^2}$$
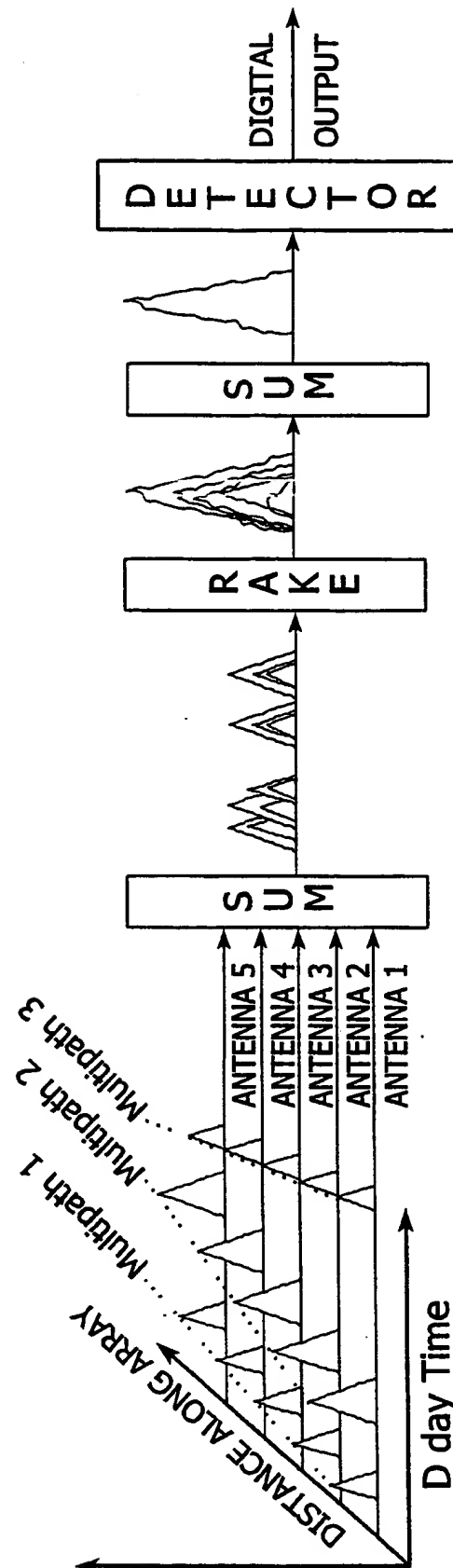
FIG. 20
PHASED RAKE PROCESSING
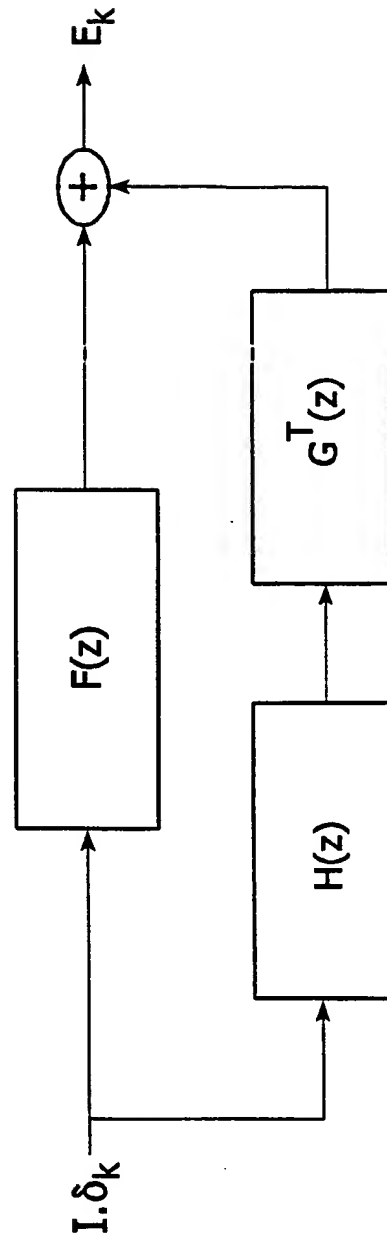
**FIG. 21**

**PRFB AS A LS MIMO FILTER DESIGN PROBLEM**

**A. CLASSIFICATION OF SUBJECT MATTER**

IPC(6) :H03K 7/06;G01S 13/00
US CL :375/44;342/75
According to International Patent Classification (IPC) or to both national classification and IPC

**B. FIELDS SEARCHED**

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 375/44;342/75

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

APS "analog (p) digital (p) pseudorandom (p) bandwidth"

**C. DOCUMENTS CONSIDERED TO BE RELEVANT**

| Category* | Citation of document, with indication, where appropriate, of the relevant passages | Relevant to claim No. |
| --- | --- | --- |
| X | US 4,893,316 A (JANE et al) 9 January 1990, col. 4 to col. 7. | 1-10, 12-38 |
| Y | | 11 |
| Y | US 4,665,401 A (GARRARD et al) 12 MAY 1987, col. 6, lines 9-40. | 11 |

☐ Further documents are listed in the continuation of Box C.   ☐ See patent family annex.

| | | |
| --- | --- | --- |
| * | Special categories of cited documents: | "T" | later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention |
| "A" | document defining the general state of the art which is not considered to be of particular relevance | |
| "E" | earlier document published on or after the international filing date | "X" | document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone |
| "L" | document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified) | "Y" | document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art |
| "O" | document referring to an oral disclosure, use, exhibition or other means | |
| "P" | document published prior to the international filing date but later than the priority date claimed | "&" | document member of the same patent family |

| Date of the actual completion of the international search | Date of mailing of the international search report |
| --- | --- |
| 08 OCTOBER 1998 | 16 DEC 1998 |
| Name and mailing address of the ISA/US<br>Commissioner of Patents and Trademarks<br>Box PCT<br>Washington, D.C. 20231 | Authorized officer<br>TIM VO |
| Facsimile No.   (703) 305-3230 | Telephone No.   (703) 308-5862 |

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